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Data-Aided Frequency-Domain 2×2 MIMO Equalizer for 112 Gbit/s PDM-QPSK Coherent Transmission Systems

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Abstract: Benefits of a low-complexity adaptive 32-tap 2×2 MIMO frequency-domain filter update by data-aided channel estimation over a time-domain filter with DD-LMS are shown. Superior stability and convergence speed is demonstrated with identical impairment tolerance.
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1. Introduction

Coherent optical demodulation in combination with digital signal processing (DSP) by means of linear filters allows equalization of all linear channel impairments of an uncompensated linear or weakly nonlinear fiber channel. The digital filter can be implemented in time-domain (TD) or in frequency-domain (FD), where FD implementation proves a lower implementation complexity for large filter memory lengths [1]. TD filtering with non-data-aided (NDA) channel acquisition by gradient algorithms like constant-modulus algorithm (CMA) or decision-directed (DD) least mean square (LMS) are widely considered [2]. As an effective NDA FD filter update is not known, data-aided (DA) channel estimation is preferred, which requires periodical transmission of a known training sequence (TS) to track time-varying polarization effects [3]. From each TS, a complete channel estimation with instantaneous filter acquisition can be obtained, which comes at the cost of additional overhead widening the spectrum of the transmitted signal. A complexity analysis and the tracking performance with constrained parallel implementation have been demonstrated in [3]. The authors of [1] and [4] discussed the widely employed dual stage equalizer with a FD chromatic dispersion (CD) compensation stage followed by an adaptive 2×2 multi-input multi-output (MIMO) equalizer. However, no performance of a 2×2 MIMO FD filter with DA channel estimation has been demonstrated.

In the following, we compare the performance of a 2×2 MIMO FD filter with a TD filter employing DA and NDA channel estimation respectively. We discuss the implementation complexity of the DA filter update, the initial convergence, the tracking speed and provide the tolerances with respect to CD, differential group delay (DGD), polarization-dependent loss (PDL) and rotation of the state of polarization (SOP) for a realistic filter design.

2. Structure of TD and FD Filter Implementation with NDA and DA Channel Estimation

For equalization and synchronization, we apply the basic DSP structure with FD CD compensation, timing recovery, adaptive 2×2 MIMO filter and carrier recovery [1, 2]. The 2×2 MIMO TD equalizer (TDE) is realized by 4 finite impulse response (FIR) filters arranged in a butterfly structure, Fig. 1-a), which mathematically relates to the de-convolution of the signal with the channel impulse response. From input samples $r_x(t)$ and $r_y(t)$ and from filtered samples $z_x(t)$ and $z_y(t)$, the CMA or DD-LMS cost function is calculated to perform a symbol-by-symbol gradient filter update (note, DD-LMS requires decision after carrier recovery). Convergence to a single source (*singularity*) is prevented by independent component analysis (ICA) added to the cost function [5]. Still, identification of the transmitted signals with respect to x- and y-polarization is required before the BER of each tributary can be evaluated. The CMA/DD-LMS update converges to the minimum-mean-square-error (MMSE) filter solution. The 2×2 MIMO FD equalizer (FDE), applies a fast Fourier transform (FFT) to overlapping blocks after serial-to-parallel conversion. For each discrete frequency component a 1-tap 2×2 filter operation is performed, which mathematically relates to the multiplication of the inverse channel transfer function, before the signal is transferred back into TD [1].

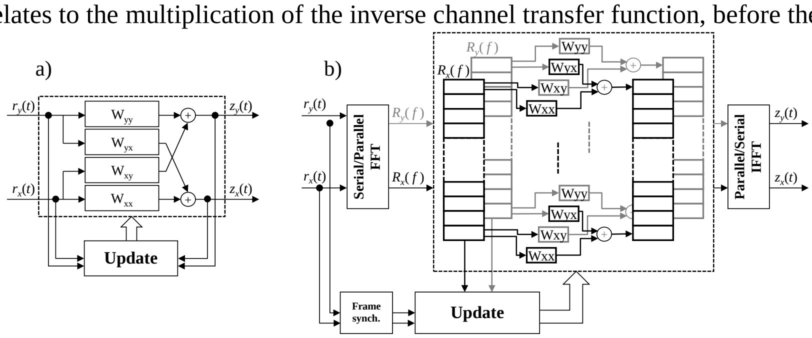

Fig. 1. Block diagram of a) NDA-TDE and b) DA-FDE.

TABLE 1
Parameter Range and Distribution for Channel Simulations

Impairment	Distribution	Value range
PMD	Maxwellian	Mean 25 ps
CD	linear	[-500: 500] ps/nm
α	linear	[0: 2π] rad
φ	linear	[0: 2π] rad

For DA filter update, a framing synchronization detects the known TS periodically transmitted in between the payload data. From the known TS $S_x(f)$ and $S_y(f)$ and the according received $R_x(f)$ and $R_y(f)$ a full channel estimation can be performed, which allows to calculate the MMSE filter transfer function, Fig. 1-b). Our TS is composed of an orthogonal constant-amplitude zero-autocorrelation (CAZAC) sequence of 16-symbol length sent in each polarization, with a 4-symbol guard interval (GI) at each side, leading to a total length of 24 symbols [3]. This allows an efficient implementation of the channel estimation for a 32 sample FFT-size. Note, no feedback and no carrier recovery are required for channel estimation, which allows avoiding processing latency and update delays. The CAZAC sequence follows a QPSK constellation. However, in principle the modulation of the TS and the payload data is independent from each other, which allows flexible switching of the data modulation format. In contrast to CMA/DD-LMS, the FD DA channel estimation can be performed on non-integer numbers of samples per symbols. Finally, the low complexity of the FD filter combined with DA channel estimation makes it highly suitable for ASIC implementation in high-speed optical receivers.

3. BER Performance Analysis

The BER performance investigation is based on simulated 112 Gbit/s polarization-division multiplexed (PDM) QPSK transmission (28 GBaud) with channel conditions including CD, all-order PMD, SOP rotation by polarization rotation angle α and polarization phase φ, PDL and additive white Gaussian noise (AWGN). For each combination of PDL values [0, 5, 10] dB and OSNR ranging between [10:1:25] dB, the BER of 100 random channels has been evaluated. The parameters for each channel realization are chosen according to the distributions in Table 1. After an optical Gaussian band-pass filter (2^{nd}-order, double-sided 35 GHz), a polarization-diverse 90°-hybrid and an electrical Bessel filter (5^{th}-order, 19 GHz), an ADC stage digitalizes the received signal at 2 samples per symbol.

The 31-tap NDA TDE, is initially converged with a symbol-by-symbol update by CMA (10^5 symbols) and then switched to DD-LMS (10^5 symbols) supported by ICA in order to avoid singularity [5]. Fig. 2-a), depicts an exemplary convergence process for selected taps. Error counting started after 2×10^5 symbols until sufficient statistics has been reached.

For the 32-tap DA-MMSE FDE, the 24-symbol TS including the GIs is repeatedly transmitted at a rate of about 10 MHz in between the payload data. Fig. 2-b) shows the required OSNR for 100 random channels at 5 dB PDL with respect to the number of channel estimation averages. Calculating the filter taps from the channel estimation of a single TS already leads to an acceptable performance with a worst-case OSNR penalty of 3 dB, while averaging over consecutive channel estimations (consecutive TSs employed) improves the filtering performance. It becomes clear that compared to NDA channel estimation with initial convergence in the range of 2×10^5 symbols, the DA acquisition is a magnitude faster in the range of 2.8×10^4 symbols.

In the following, we compare the BER performance of the DA-MMSE FDE with filter update averaging over 10 channel estimations to DD-LMS TDE. For each combination of PDL and OSNR values, Fig. 2-c) and -d) show the minimum, the mean, the standard deviation around the mean and the maximum BER performance over 100 random channel trials within the parameter range from Table 1. Despite a low 0.5 dB OSNR penalty, which we expect to be overcome in future with enhanced DA channel estimation, the mean BER of the DA-MMSE FDE shows almost identical behavior as the DD-LMS TDE. This confirms that both methods converge to the MMSE solution. However, it should be noted that the standard deviation of the BER for the DD-LMS TDE is much larger with a wider spread of minimum and maximum BER. This results from the NDA gradient filter update, which allows sub-optimum convergence. For strong 10 dB PDL, the initial channel acquisition with ICA [5] could not be finished

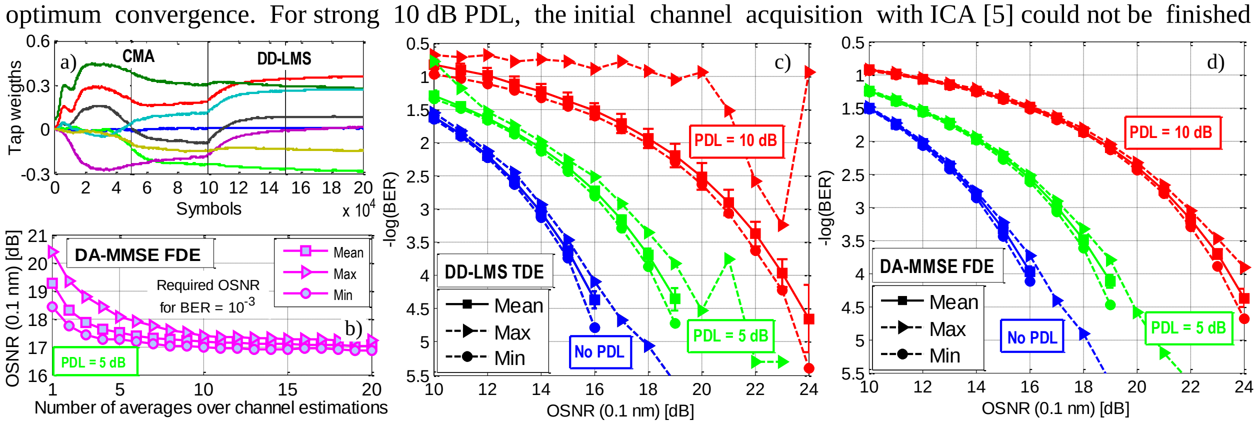

Fig. 2. DD-LMS TDE and DA-MMSE FDE performance comparison. a) DD-LMS TDE convergence of selected filter taps. b) DA-MMSE FDE OSNR required at BER=10^{-3} for different channel estimate averages. BER measurement randomly varying CD, DGD, SOP rotation (parameter distributions presented in Table 1) for each PDL value [0, 5, 10] dB, for c) DD-LMS TDE and d) DA-MMSE FDE.

within the given 2×10^5 symbols, which leads to failure in about 3% of the channel realizations. In contrast, DA-MMSE FDE prevents convergence to suboptimum filter solutions and instabilities such that no ICA and no polarization identification are required.

4. Impairment Tolerance of DA-MMSE FDE

Given the same configuration as in the previous section, we investigated the impairment tolerance of the DA-MMSE FDE with respect to CD, DGD, PDL and SOP rotation. The filter taps are based on 10 channel estimate averages.

In Fig. 3-a), the CD tolerance proves penalty-free performance up to 1000 ps/nm, which corresponds to channel memory of 7.8 symbols. This clearly indicates the importance of the GI used at each side of the TS, which covers a total of 8 symbols. For higher values of CD, the GIs cannot entirely absorb the pulse spread and consequently we get BER degradation. We observe a smooth degradation with a 1 dB OSNR penalty around 1500 ps/nm for BER=10^{-3}, allowing transmission over dispersion managed links or short-range metro networks without FD CD compensation.

Similarly, penalty-free DGD tolerance presented in Fig. 3-b) covers 260 ps, which again reflects the 8-symbol channel memory covered by the GIs. In contrast to the CD tolerance, higher values of DGD experience a steep degradation. Large DGD tolerance is typically not required, making it clear that the GI has been mainly designed to meet the CD requirement, which follows the worst-case remaining CD after the first CD compensation stage [4].

Despite the channel impairments with memory, PDL is one of the limiting memory-less effects. It results in a polarization-dependent OSNR degradation which can be estimated but not compensated by filtering. We assume *worst-case* PDL attenuating the signal of only one polarization, which leads to a *worst-case* BER performance. In Fig. 3-c), the required OSNR follows the same tolerance as TDEs with CMA/DD-LMS update with 1 dB OSNR penalty at 3 dB PDL and 3 dB OSNR penalty at 5.7 dB PDL [6].

The time-variant channel characteristic is mainly caused by SOP rotation. Due to the large degree of parallelization in the DSP architecture and due to processing latency, the filter update might be delayed. Therefore, the SOP-detuning tolerance is vital to estimate the penalty for implementation-constrained channel acquisition and to design the required repetition rate of the TS. In Fig. 3-d), the filter taps are estimated for α=0 and φ=0. Detuning the polarization angle by $\Delta\alpha$, a gradual degradation is observed with 1 dB OSNR penalty at $\Delta\alpha$=9° for BER=10^{-3}. For SOP rotations of 20 kHz an update rate in the range of 10 MHz seems to be appropriate, which results in less than 2% overhead for the given TS including the GIs.

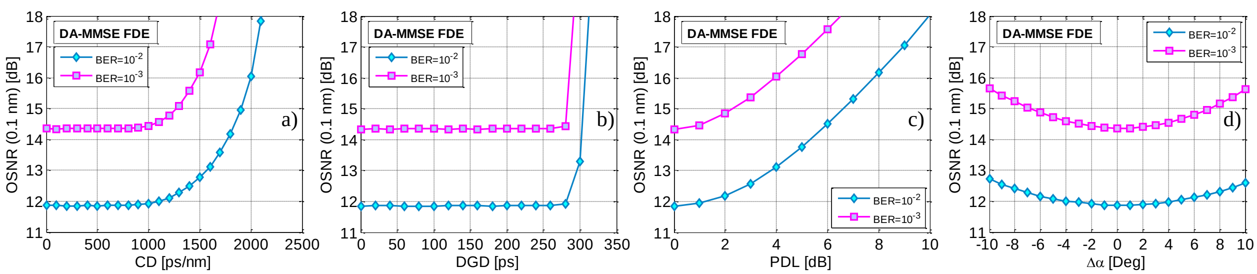

Fig. 3. Impairment tolerance of DA-MMSE FDE with respect to a) CD, b) DGD, c) PDL and d) SOP rotation.

5. Conclusions

BER performance based on DA-MMSE FD channel equalization has been shown in single carrier coherent receivers. With a similar BER performance compared to the state of the art DD-LMS TD equalizer, the FD equalizer exhibits optimum and faster convergence using overhead for training below 2%. Furthermore, the low complexity architecture and the feed-forward DA filter update make the FD equalizer highly suitable for implementation in high-speed optical receivers.

6. Acknowledgement

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